A 600 MEGACYCLE CONVERTER FOR USE WITH PULSE-TIME TELEMETRY AND RADAR RANGING RECEIVERS

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A 600 MEGACYCLE CONVERTER FOR USE
WITH PULSE-TIME TELEMETRY AND
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PREFACE

The design and development of any piece of equipment involves a large amount of investigation and experimental work. This thesis is a treatise on the design of a radio-frequency converter unit which is to be used for a specific purpose. The background details, basic theory, design problems, and results are explained.

I wish to thank Dr. H. L. Jones for his interest and valuable criticism in the preparation of this thesis. I also wish to thank the staff of the Research Foundation, Electronics Laboratory for their help in the development project. The availability of the facilities of the Electronics Laboratory are also appreciated.

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CHAPTER I

INTRODUCTION

During recent years the Research Foundation of Oklahoma A. and M. College has undertaken to develop and produce small production quantities of radar beacons. These beacons, sometimes called transponders, consist essentially of a radio-frequency receiver and transmitter. The beacon is a link of a radar triangulation system which utilizes the propogation time of pulsed radio-frequency signals to plot trajectory information of airborne vehicles. Other uses of the beacons are to perform command functions within the vehicle and to respond with information relative to its environment. The latter function is called telemetry. The beacons are compact, light weight, and ruggedly constructed. Power consumption is kept to a minimum.

The receiving portion of the beacon consists of a mixer-oscillator or converter unit and a four-stage intermediate-frequency amplifier with a diode detector. The receiver has a pulse operated automatic gain control circuit. The intermediate-frequency amplifier consists of two staggered pairs of amplifiers with an overall bandwidth of approximately three megacycles and an overall voltage gain of about 80 decibels. The signal required for a 2/1 signal/noise ratio is about 35 microvolts with a detector output noise level of approximately one-tenth of a volt.

Like many other types of mechanical or electrical devices the beacon and its component circuitry undergo certain design modifications each time a new quantity is produced. These modifications are made to conform to requirements of the purchaser and also to take advantage of newly developed and improved components. One unit of the beacon which has had but little improvement is the radio-frequency converter.

The converter is a unit which in general performs satisfactorily but usually requires a certain amount of haphazard adjustment. The design of the unit is one that does not lend itself to duplication where performance is concerned. The tuning elements for the oscillator and the mixer are small parallel transmission lines tuned with adjustable shorting bars. The mixer line has a variable capacitor for additional tuning. The oscillator tube is a subminiature type 5703 triode and the mixer tube is a subminiature type 5702 pentode which is triode connected. Both of these tubes have lead inductances which are high because of the small size of the leads. These lead inductances are subject to some variation due to lead length and arrangement in wiring the unit. The interelectrode capacitances of the tubes are also subject to some variation which alter the performance of the circuits in general. It is known that the mixer tube is operating very close to internal resonance at 600 megacycles. With slight deviations of inductance or capacitance the receiver sensitivity loss due to mistuning is sometimes difficult to cope with. There have been instances when a mixer tube has been replaced with another tube to improve conversion gain when for most purposes one tube was as good as another.

For these and other reasons it has been thought desirable to have a new design for the converter unit for some time. The design will be described in this thesis.

Since the frequency of operation of the beacon is 600 megacycles the tuning elements may be distributed transmission line elements or a combination of distributed and lumped elements. It has been decided that any new design for the converter unit would be one which utilized distributed transmission line tuning elements. These tuning elements may have been coaxial transmission lines, flat plate lines or parallel lines. It was also desired to use vacuum tubes which were designed for use at this frequency and were adaptable to some available tuned-line configuration without the use of any connecting leads. Any improvement in overall sensitivity for the beacon receiver would be readily accepted and it was hoped that a two-to-one improvement could be obtained. It was also anticipated that the selectivity of the converter unit could be improved considerably. This would help in preventing the transmitter output signal from spilling into the receiver. Transmitter spill-through has been a problem in the past because a common 50 ohm antenna system has been used for transmitting and receiving. The beacon transmitter frequency is 580 megacycles. This is only 20 megacycles displaced from the signal input frequency, consequently the selectivity would need to be quite sharp.

The goal for size, weight and power requirements were set at nothing larger than that of the old converter. The old unit measures 2 x 3 x 4 inches and weighs 3/4 lbs. It draws 23 milliamperes at 150 volts and 0.4 amperes at 6.3 volts. The frequency tuning range of the new unit was to be for an incoming signal frequency spread of 590 to 610 megacycles. In order that the unit would meet most any environmental specification which might be imposed it was to be designed to meet the following requirements: The frequency of the local oscillator should not shift more than one megacycle in either direction nor should the sensitivity of the whole receiver suffer more than a two-to-one change due to a change of environment of the converter unit when subjected to the following conditions: The temperature of environment should be allowed to change from -55°F to +160°F. The unit should withstand three scans of vibration from 10 to 55 cycles per second with a double amplitude of 0.032 inches in each of its three major planes. It should withstand three impact shocks in both directions of each of its major axes. Each impact should have 15 g acceleration for a duration of 0.011 seconds. No humidity, moisture, or altitude conditions were imposed since it was expected that the unit would normally be operated within a pressurized dry container.

CHAPTER II

DISCUSSION OF CIRCUIT CONSIDERATIONS

A superheterodyne receiver employs the technique of combining the incoming modulated radio-frequency signal with a locally generated continuous-wave signal. They are combined in some type of nonlinear device and consequently, a difference frequency is obtained. This difference frequency, called the intermediate frequency, is further amplified in the intermediate-frequency amplifier. The local oscillator and the mixer are in combination referred to as the converter.[1] The converter unit may also incorporate one or more stages of radio-frequency amplification and/or passive radio-frequency circuits. From these facts it is obvious that an oscillator, mixer, and possibly, an amplifier and a passive preselector will be employed in the design of a converter unit. These will involve the use of vacuum tubes and tuned circuits. The important point here is to find the type of each which is most adaptable to this problem.

From previous experience it has been fairly well established that the tuning elements for this converter will need to be of the transmission line type of circuit with little if any added lumped reactances. It is known that sections of transmission lines which

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[1]Terman, F. E., Radio Engineers Handbook, McGraw-Hill 1953, Page 567.

are multiples of a quarter wavelength at some frequency may be made to act like resonant or antiresonant circuits at that frequency, if the sections are terminated with a short or open circuit. A desired reactance can also be obtained by adjusting the length of a short-circuited or open-circuited section of transmission line. There are various configurations of transmission lines which could be used for radio-frequency tuned circuits. Among these are coaxial lines, parallel lines, and flat plate lines. The coaxial line seems to carry more advantages for usage here than either of the other two. Primarily it has had ultra-high-frequency vacuum tubes designed for its use. These tubes are the familiar disc seal "lighthouse" coplanar triodes, rocket tubes, and the newer pencil triodes. There are few tubes which are adapted to the flat plate or the parallel lines at these frequencies except by makeshift methods.[2] The coaxial line circuits have another advantage in that they provide their own shielding. This is useful for reducing radiation from the local oscillator, and reducing reception of extraneous signals in the radio-frequency amplifier circuits. The chief disadvantage of using coaxial circuits is that their construction is more complicated than any of the others mentioned.

As far as vacuum tubes are concerned the types mentioned above are required for coaxial-line circuits. Miniature and subminiature tubes have been ruled out because of their construction. This leaves us with the disc seal tubes and coaxial elements. The pencil triode has been chosen for its small size and reasonable price.

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[2]Radio Research Laboratory Staff. Very High Frequency Techniques. Vol. I, McGraw-Hill 1947, Page 337.

Other circuit elements are the bypass capacitors and bias resistors. The resistors are composition resistors of the common variety. All bypass capacitors are constructed within the unit wherever possible. Capacitors made of flat plates separated by a dielectric are generally better than commercial units with wire leads when used at these frequencies. Dielectric material used will be tetrafluorethylene, known as teflon. This material is an excellent dielectric for use at ultra-high frequencies.

The basic circuit for the converter unit is the oscillator for generating the local signal and some type of mixer for frequency conversion. The converter described here will consist of three coaxial line circuits and a crystal mixer. The oscillator and a radio-frequency amplifier will be of the folded-back common-grid or grid-separation type of coaxial circuit using pencil triodes. The crystal type mixer will be used because of its low noise characteristics.[3] A passive coaxial preselector will be used between the antenna and the radio-frequency amplifier. The purpose of the preselector is to improve the selectivity of the converter and to help facilitate the use of a common antenna for the beacon. As previously stated, the transmitter spill-through has been a problem. The selectivity of the preselector will help reject the powerful signal of the transmitter by presenting a high impedance to the transmitter frequency. This high impedance may also be utilized to more effectively direct the transmitter power to the antenna when a common antenna system is used. The amplifier is

[3]Radio Research Laboratory Staff. Very High Frequency Techniques. Vol. II, McGraw-Hill 1947, Page 798.

used to overcome conversion loss in the crystal and signal loss in the preselector and to improve on the overall receiver sensitivity as much as possible.

It is not the purpose of this thesis project to develop any radically new circuits but to adapt already existing design information to this particular converter unit. It is known that efficient oscillators and amplifiers of the coaxial variety are in existence and have been used in radar and other ultra-high-frequency circuits quite extensively. This thesis purports to show how these circuits can be made to perform together as a unit to accomplish what has been set forth. No detailed theory of operation will be given but a general explanation of circuit operation will be made.

CHAPTER III

CIRCUIT THEORY AND DESIGN

The three major units of the converter are the oscillator, the amplifier, and the preselector. The fourth unit, the crystal mount, is attached to the amplifier unit. It was known from the beginning that all of the units would eventually be attached together as one complete converter. This involves coupling energy from one tuned cavity into another as well as other intricate mechanical problems which are sometimes difficult to solve. During the first stages of design the oscillator, amplifier, and preselector were treated as individual units. The oscillator, for example, was the first unit designed and was given full attention until its status was advanced enough for the amplifier to be designed. After the amplifier was completed and was performing satisfactorily its output connector was replaced with the crystal mixer. The local-oscillator signal was coupled into the amplifier by means of coaxial cable, and the three were made to function as a converter. Finally a preselector was built, checked out as an individual unit, and placed in the input circuit with coaxial cable. Its usefulness was verified, however, it did not contribute much to overall performance except that of selectivity.

As previously mentioned the oscillator and amplifier use the grid-separation type of circuit. This circuit is demanded because of the configuration of the pencil triode tubes and the mechanical methods of attaching the coaxial lines to the tubes. In grid-

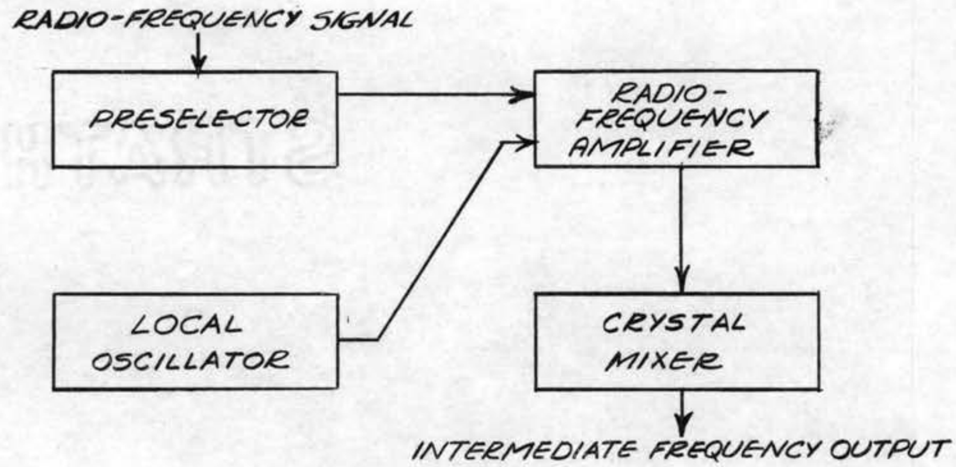

Figure 1. Block diagram.

separation circuits the input voltage is applied between cathode and grid and the output is coupled from the grid-plate resonator rather than from the plate-cathode resonator as in the usual circuit.[4] Figure 2 shows schematically the voltages and currents

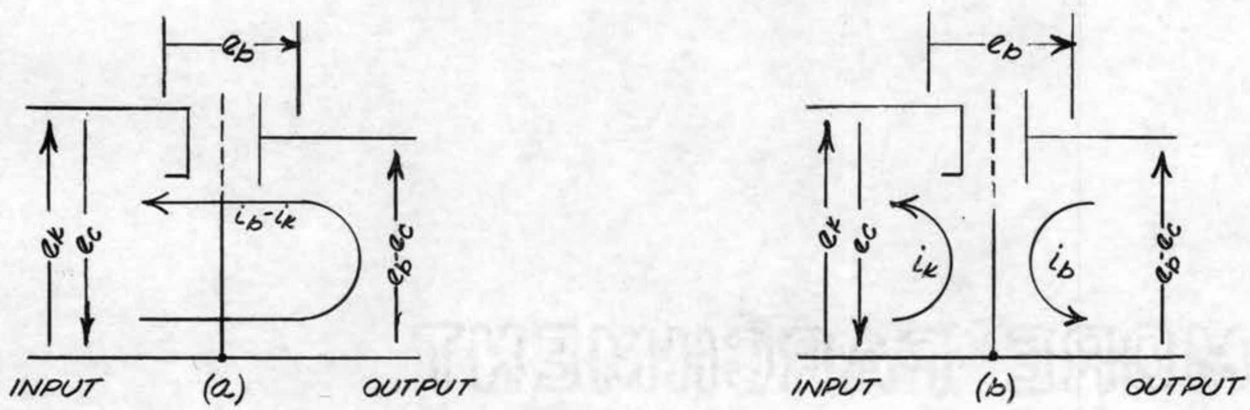

Figure 2. Grid separation circuit.

in the grid-separation or grid-return circuit. When the signal levels are low there is generally enough bias to prevent grid current

[4]Reintjes, J. F. and G. T. Coate. Principles of Radar. McGraw Hill 1952, Page 716.

flow and Figure 2a applies. The input signal voltage is e_k or $-e_c$. The grid-plate voltage, $e_b - e_c$ is the output voltage. The plate current and the cathode current are equal. This current is the negative of the current supplied to the cathode from the signal source or the current supplied by the plate to the load. In case there is grid current in the circuit, the plate current does not equal the cathode current and Figure 2b is used. If the tube has a large amplification factor the electric field between grid and plate do not influence the electrons between grid and cathode. The grid and cathode act largely as a diode with the cathode current, i_k being determined by the signal voltage, e_k. Most of the electrons reaching the grid pass on to the plate. The electrons are accelerated to the plate by the direct voltage between grid and plate. They are decelerated by the alternating component of voltage on the plate, thus the direct current energy of the power supply is converted into alternating current energy to be stored in the grid-plate resonator. The resonators in this instance are tuned sections of coaxial transmission line.

A section of lossless transmission line which is terminated with a short circuit appears as a reactance when viewed from the open end. This reactance is inductive, capacitive, infinite, or zero depending on the length of the line. The value of the reactance is represented by

$$X = jZ_0 \tan \frac{2\pi \ell}{\lambda}$$

where Z_0 is the characteristic impedance of the transmission line, ℓ is the length and λ is the wavelength within the dielectric for the frequency concerned.

As illustrated in Figure 3, the input reactance is inductive for any section of short-circuited transmission line which is less than a quarter wavelength long. From this fact it can be seen

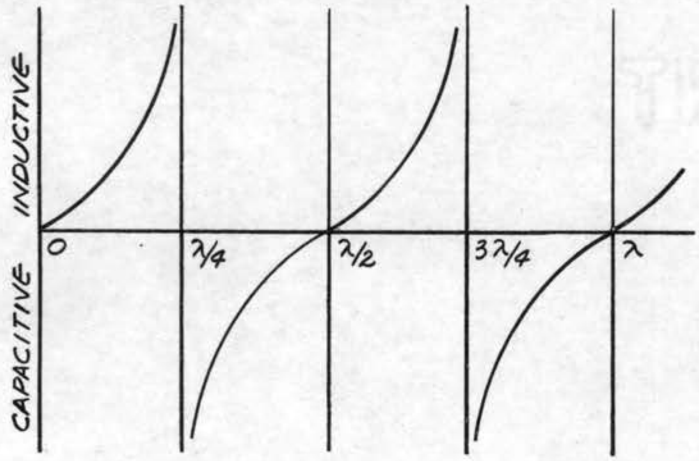

Figure 3. Input reactance vs. wavelength for short-circuited transmission line.

that any capacitive reactance may be tuned to resonance by paralleling it with a short-circuited transmission line of equal inductive reactance. This is an example of the tuned circuits which are used in the cavities of the grid-separation circuits of this converter. As an example, the grid-plate capacity of the 5675 oscillator tube is 1.3 micromicrofarads. This represents 194 ohms of capacitive reactance at 630 megacycles. A matching inductive reactance using a 70 ohm shorted line will require a line length of 9.4 centimeters. Actually it is generally found that other factors enter into the calculations and the final length of line is obtained by experiment.

The characteristic impedance, Z_0 of a transmission line is given by

$$Z_0 = 138 \log (b/a)$$

where b and a are the outside and inside radii respectively.

The maximum Q for a coaxial resonator occurs when the characteristic impedance is 75 ohms. It will not be reduced more than 15 percent for impedances of 42 or 121 ohms. The unloaded Q is given by the expression below where (b) and (a) are measured in centimeters.

$$Q = \frac{0.694 \sqrt{\text{freq.}}\ (b) \log (b/a)}{1 + b/a}$$

In designing the coaxial cavities for the oscillator and amplifier the size of the brass tubing used for the grid cylinder is determined by the diameter of the grid disc on the pencil triode. The innermost rod, called the plate line, and the outermost cathode cylinder are chosen in the design to satisfy design requirements. The plate line was made with a 0.250 inch outside diameter. This is the same diameter as the anode connector of the pencil triode. The small size of the plate line helps reduce weight and also aids in accomplishing dielectric tuning which will be explained later. The grid-plate line characteristic impedance with a 0.875 inch grid sleeve and a 0.250 inch plate line is 70 ohms. This gives maximum cavity Q which is desirable in the continuous-wave local oscillator for better frequency stability. The outer sleeve diameter was chosen as 1.250 inches because this is about as small as could be used without approaching the diameter of the grid sleeve. This gives a characteristic impedance for the grid-cathode cavity of 18 ohms. It was felt that since the grid-cathode conductance in grid separation circuits is high, the low line impedance would not greatly affect the circuit operation.

Power can be coupled from a coaxial cavity by either of two methods. One is the loop method whereby a loop of wire is introduced near the shorted end of the grid-plate cavity. This places the loop where the radio-frequency current in the cavity is highest and it is oriented so that it will couple energy from the magnetic field. This places it in a plane perpendicular to the circumferential flux lines. In some cases it may be rotated to vary the coupling. The other method is called capacitance probe coupling. This uses a small disc which is entered into the grid-plate cavity at a point of maximum electric field intensity. In the quarter wavelength cavity the point of maximum intensity is near the tube anode. The probe should be adjustable radially to adjust its capacity with respect to the plate line and consequently its coupling efficiency. Power coupling from the local oscillator of the converter is not a problem of obtaining efficiency but more a matter of doing it in the most convenient manner. The power output required is trivial. It was decided that in this design loops would be used since they are easier to build.

Local Oscillator

Since the local oscillator was designed prior to any of the other units its treatment will be taken up first even though the amplifier is a little more basic. The general layout configuration for oscillators is the so called "end-to-end" oscillator shown in Figure 4. The resonant cavities are actually short-circuited

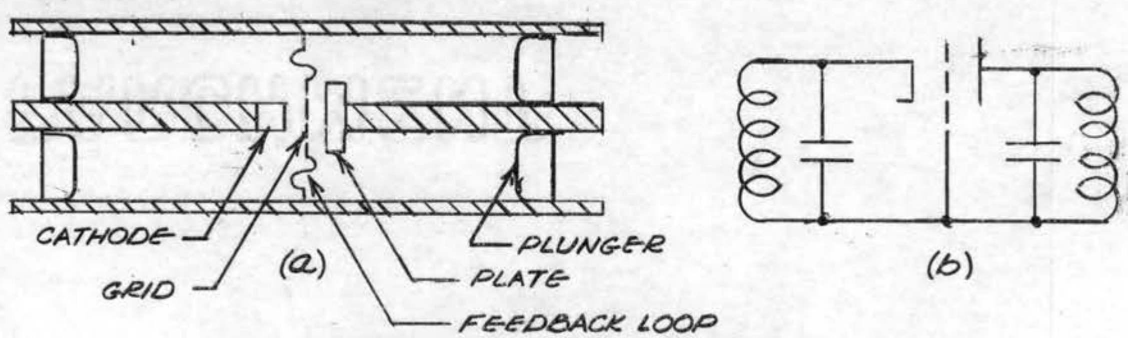

Figure 4. "End-to-End" oscillator.

sections of coaxial transmission line which are approximately one quarter of a wavelength long. They are shunted on their open ends by the tube capacities. Schematically they may be represented by Figure 4b. As illustrated, this circuit has the grid-cathode cavity on one side and the grid-plate cavity on the other. Both cavities are adjustable with movable metal plungers. The only means of coupling energy between the two resonator cavities is through the electron beam and by electromagnetic coupling through the grid.[5] In the amplifier the electromagnetic coupling is desired to be a minimum to avoid regeneration but for the oscillator, feedback must be added in order to sustain oscillation. Various forms of loops, capacity probes and combinations of both are used for feedback as indicated in Figure 4. The feedback energy must satisfy certain phase and amplitude requirements to sustain oscillation.

The "folded-back" oscillator is electrically similar to the "end-to-end" oscillator. It is derived from the "end-to-end" oscillator by folding the grid-cathode cavity over the grid-plate cavity. This is illustrated in Figure 5. A suitable feedback arrangement is shown with the threaded feedback screw. This is a capacity coupling probe which couples energy from the grid-plate cavity into the grid-cathode cavity. When the tuning plungers are adjusted for maximum efficiency with proper feedback adjustment, it will generally be found that the grid-plate cavity is electrically longer. The advantages of the folded-back design are in its small physical length and ease of tuning. The small size is the reason

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[5]Gurewitsch, A. M. and J. R. Whinery. "Microwave Oscillators using Disc-Seal Tubes." Proceedings of IRE, Vol. 35, No. 5, May, 1947.

for using this design on this project.

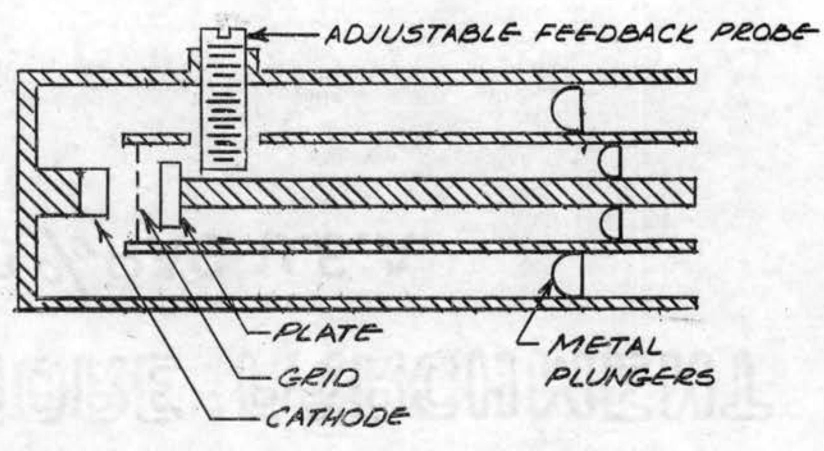

Figure 5. "Folded-Back" oscillator.

To analyze the electrical performance of the oscillator refer to Figure 6.[6] L_p and L_k represent the inductances of the grid-plate and grid-cathode coaxial cavities repsectively. C_{gp}, C_{gk} and C_{pk} represent interelectrode and stray capacities. C_{pk} is augmented by the feedback capacity screw as shown in Figure 5. The feedback voltage to sustain oscillation is applied to the cathode by the capacity voltage divider as indicated in Figure 6b. C_k is the resultant capacity obtained when the grid-cathode cavity, with C_{gk} has a resonant frequency lower than f_o, the frequency of oscillation. This makes the cathode cavity appear capacitive at f_o.

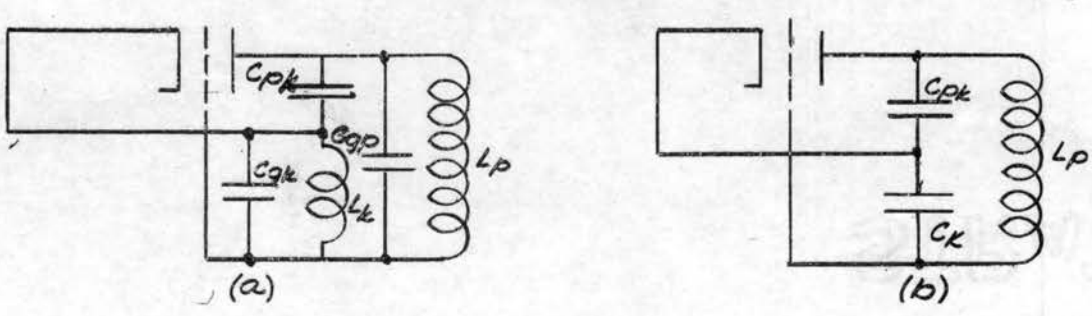

Figure 6. Radio-frequency equivalent circuit of oscillator.

[6]Reintjes, J. F. and G. T. Coate. Principles of Radar, McGraw-Hill 1952, Page 721.

The grid-plate cavity resonates with C_{gp} at a frequency higher than f_o making it an inductive reactance. This inductive reactance, L_P will resonate with the series combination of C_{pk} and C_K at the oscillator frequency, f_o.

A cross-section view of the oscillator, amplifier and preselector is shown in Figure 7. These cavities do not use adjustable tuning plungers but use other tuning methods to be described later. It may be noted that the overall length of the outer cavities is four inches. This length was determined when the oscillator was designed and the other units were forced to comply in length. The four-inch length was established by the length of the grid-plate cavity of the oscillator and the additional length required to accomodate the pencil triode. The gird-plate cavity was made 3.4 inches long which is correct to produce a local-oscillator frequency of 630 megacycles. This frequency is 30 megacycles above the signal frequency of 600 megacycles and consequently gives a difference frequency of 30 megacycles which is the frequency passed by the intermediate-frequency amplifier.

The grid-plate cavity is electrically lengthened and its resonant frequency lowered by a teflon disc near the tube end. This disc is used primarily to support the plate line within the grid cylinder. Its thickness may be varied as a means of setting the mean frequency of the oscillator. The frequency is also lowered by a traveling dielectric tuning plunger within the grid-plate cavity. This plunger almost fills the area between the plate line and grid cylinder and is one inch long. It has a total travel of one inch and is positioned with an adjustable nylon screw as indicated in Figure 7. The tuning plunger is made of polystyrene base plastic called "Polypenco Q 200.5" and manufactured by The Polymer Corporation of Pennsylvania. The dielectric constant of this material is about 2.5 at 600 megacycles and its volume resistivity is greater than 10^{15} ohms per cubic centimeter. It is stable over a wide temperature range. The characteristic impedance of the line is reduced over the length of the plunger and the resultant input reactance of the cavity can be made to vary with plunger position. It lowers the frequency as it is moved near the end of the line. The total range of the oscillator frequency is 622 to 638 megacycles with this tuning plunger. The dielectric plunger is a smooth operating, noiseless method of frequency control. The original plungers used in the design were made of a tough nylon material called "Polypenco FM 10001." It performed quite satisfactorily at room temperature but has a high loss factor at higher temperatures.

As can be seen from Figures 7 and 8, the plate line does not attach directly to the end plate of the cavity but is insulated and carried through the end plate to the power supply. The plate line is bypassed for radio-frequency current by the built-in disc capacitor, C107. This capacitor consists of two brass discs insulated from the end plate with 0.020 inch thick teflon sheet. It has a calculated capacity of 40 micromicrofarads and a reactance of 7 ohms at 630 megacycles.

The grid-cathode cavity of the oscillator requires an electrical length greater than its actual "air" dimensions in order to oscillate.

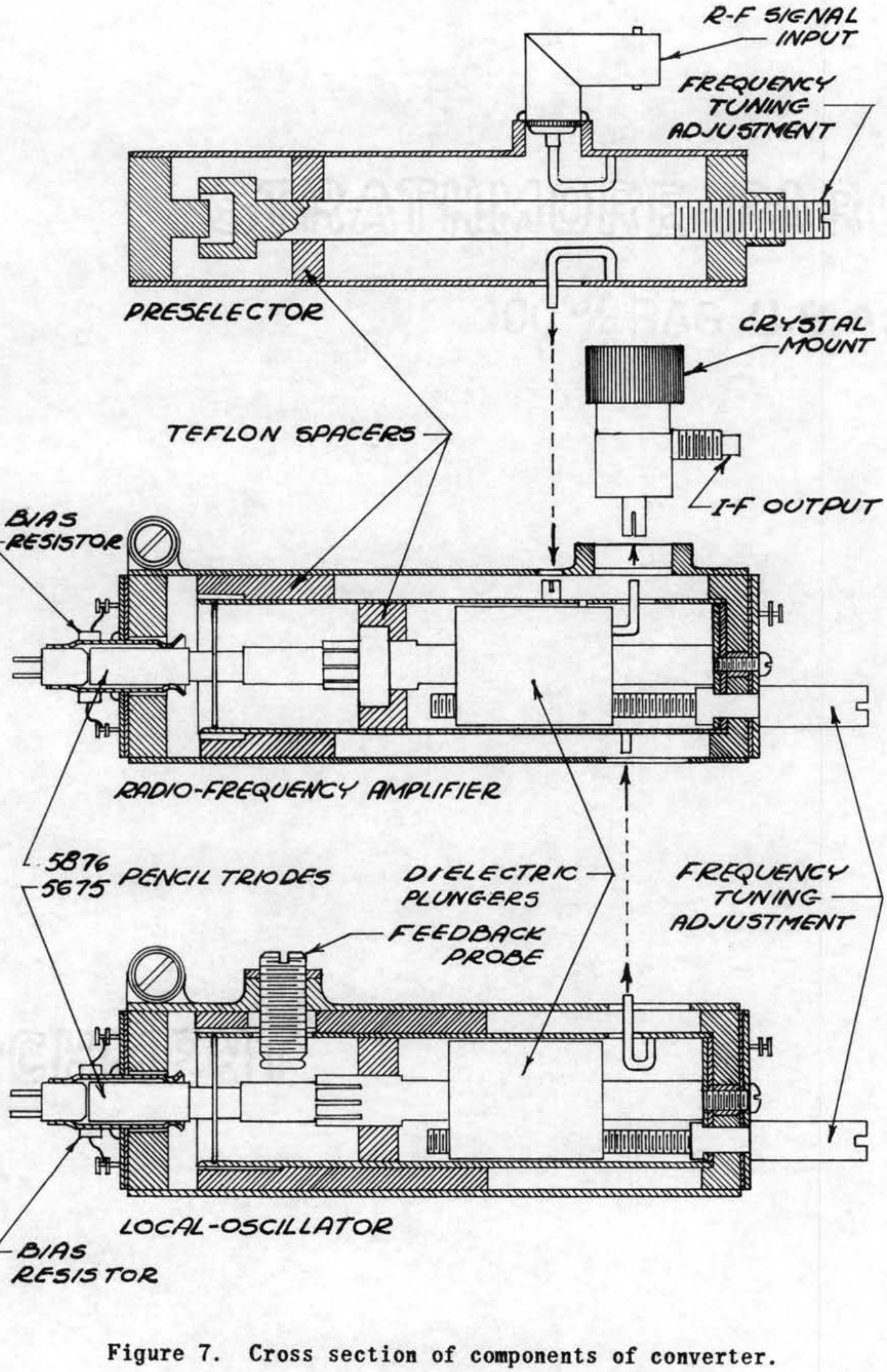

Figure 7. Cross section of components of converter.

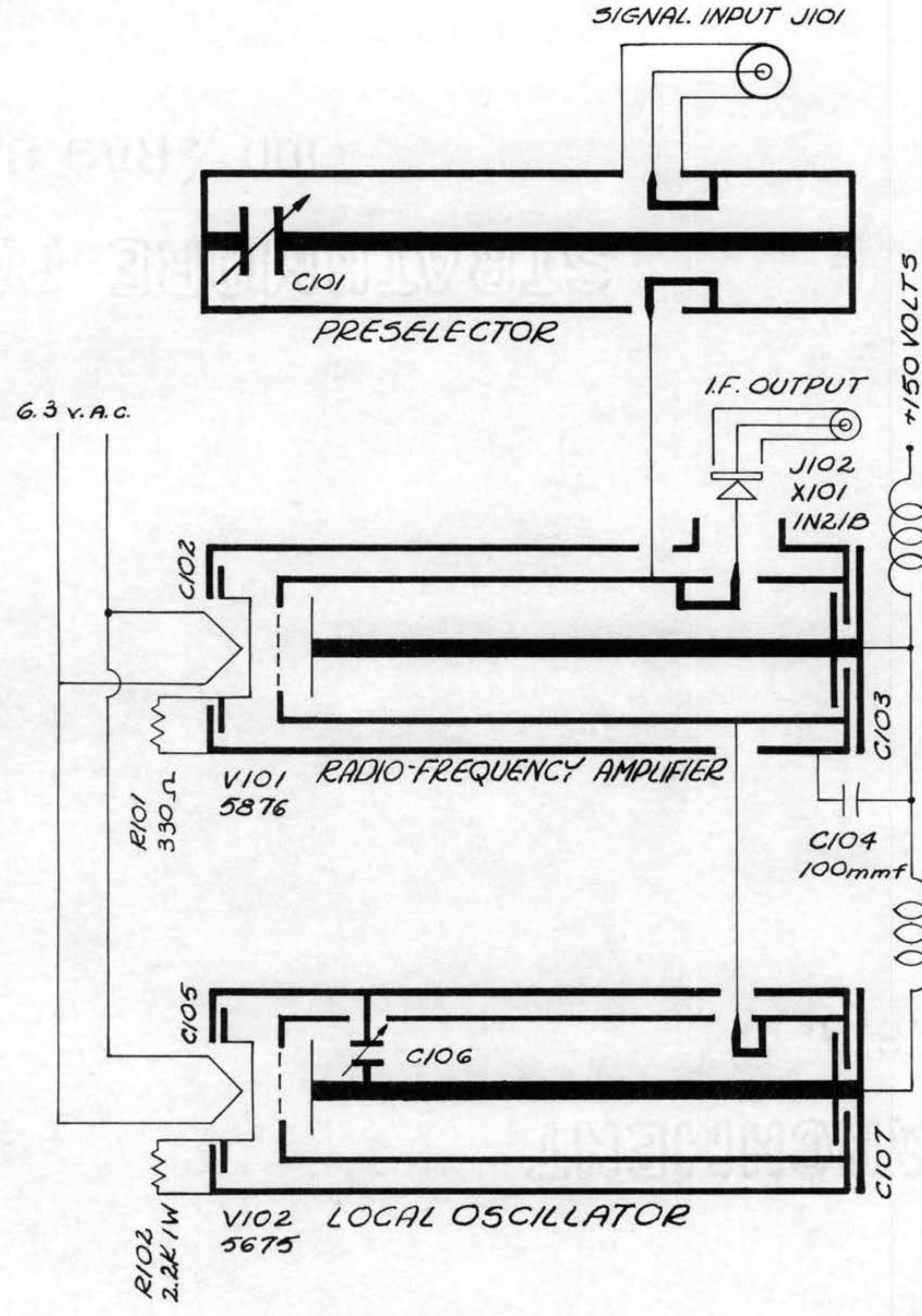

Figure 8. Schematic diagram of converter unit.

This is accomplished by the insertion of a 1.63 inch long teflon cylinder between the grid and cathode cyclinders. The location of the teflon cylinder is shown in Figure 7. The length of the cylinder was established empirically for proper feedback phase for oscillation. The reasons for this adjustment have been explained previously.

Although grid bias might be more desirable in a class C oscillator the grid cylinder may not conveniently be insulated from ground in this unit. The oscillator uses cathode bias as seen in Figure 8. The cathode of the tube is connected to a capacitor disc by means of beryllium copper contacting fingers. This disc is part of a bypass capacitor, C 105, similar to that in the plate circuit. It has a capacity of 30 micromicrofarads and a reactance of 9 ohms at 630 megacycles. The cathode bias resistor, R 102, is connected between a terminal on the disc and ground. The value of the resistance is 2200 ohms. This value is higher than would normally be used in the circuit but it reduces the strength of oscillations and reduces the power requirements from the power supply. With this resistance the oscillator was found to quench if the bypass capacity was greater than 50 micromicrofarads. Quenching is a periodic interruption of the oscillation due to a buildup of bias across the bias resistor and capacitor. Reduction of the bias time-constant is necessary for reduction of the tendency for quenching.

Feedback for the oscillator is the same as described previously for the "folded-back" oscillator. A 5/16 brass screw probe is adjusted for the optimum feedback capacity, C 106, between the cathode and plate lines. Adjustment of the feedback probe has a direct effect upon the frequency of oscillation since the feedback capacity is one of the frequency determining elements. The fixed constants of the oscillator have been selected to allow the optimum feedback adjustment to occur when the frequency is centered approximately at 630 megacycles. The frequency may be adjusted slightly with the feedback probe without encumbering the performance of the oscillator.

The type 5675 pencil triode is mounted within the cavity by insertion of the anode terminal into the plate line gripping fingers and allowing the grid disc to snap into a circular groove near the end of the grid cylinder. The grid cylinder is slotted into fingers near the end to expand over the grid disc. The cathode plug is a separate assembly which contains the cathode connecting fingers, bias resistor, and bypass capacitor. It is installed over the cathode of the tube and within the outer cylinder of the oscillator. The structural design has been made to relieve mechanical strain on the tube as much as possible.

Power is coupled out of the oscillator by means of a very small loop located one-half inch from the shorted end of the grid-plate cavity. The power supplied by the local oscillator is very small, being only about 0.5 milliwatt.

Radio-Frequency Amplifier

Many of the details that were mentioned about the design of the local oscillator also apply to the radio-frequency amplifier. As shown in Figure 7, the amplifier was made to be consistent with the oscillator in that the resonator elements, dielectric tuning plunger, the built-in bypass capacitors, and the cathode plug assembly are similar to those of the oscillator. The amplifier lacks the feedback probe which is necessary in the oscillator. The cathode bias resistor sets the plate current at 5 milliamperes with +150 volt plate supply.

The anode connector of the plate line has a 0.560 diameter by 0.200 thick brass ring attached. This ring has a cup-shaped teflon bushing about it to maintain concentricity for the grid-plate cavity. The purpose of the brass ring is to add a lumped amount of capacity across the grid-plate cavity and thereby reduce its resonant frequency to a mean value of 600 megacycles. Without this added capacity the frequency of the grid-plate resonator would have been 625 megacycles. With the mean frequency of 600 megacycles the dielectric tuning plunger allows a tuning spread of 592 to 608 megacycles which is about right to track the oscillator for a 30 megacycle difference frequency.

A 0.88 inch long teflon sleeve is located between the grid cylinder and the outer cylinder to lower the resonant frequency of the grid-cathode cavity to near 600 megacycles. The grid-cathode tuning is very broad because of the low input impedance of the amplifier.

Because the input impedance is in the order of only a few hundred ohms the problem of matching the input to a 50 ohm source is much simplified. The signal input is connected to the grid cylinder about one-third the distance from the shorted end to the grid. This is a compromise location with the preselector as explained later. This location gives an input standing wave ratio of about 2.3/1 at 600 megacycles. The local-oscillator output signal is also connected to the grid cylinder one-half inch from the shorted end of the cavity. The local-oscillator signal is injected into the crystal mixer after being passed through the amplifier.

The output coupling circuit of the radio-frequency amplifier is a 0.5 by 0.2 inch coupling loop located 0.75 inch from the short-circuited end of the cavity. The loop passes through the grid-cathode cavity into the crystal mount. The loop couples the 600 megacycle signal energy and energy from the local oscillator into the crystal mixer where they are converted into a 30 megacycle intermediate-frequency signal. The loop provides just enough loading to give the cavity a loaded Q of 100 and a bandwidth of 6 megacycles.

It is difficult and impractical to obtain some of the necessary information for making accurate calculations for such things as gain, bandwidth, and input impedance for an amplifier of this type. To be strictly accurate the calculations would contain figures for transit time loading and an accurate plate load resistance figure. A straight forward calculation for stage gain and grid-cathode input impedance is given below.

The Q of the amplifier grid-plate cavity as found from bandwidth measurements is

$$Q = \frac{f}{\Delta f} = \frac{600}{6} = 100$$

The resonant impedance is considered the plate load for the tube and is[7]

$$Z_s = \frac{4Z_o}{\pi} Q = \frac{4(70)(100)}{\pi} = 8{,}900 \text{ ohms}$$

[7]Terman, F. E., "Resonant Lines in Radio Circuits," Electrical Engineering, Vol. 53, No. 7, July 1934.

where the characteristic impedance, Z_0 is considered to be 70 ohms even though it may be modified by the dielectric tuning plunger.

The gain of the tube from input to output is given by[8]

$$\text{Gain} = \frac{(\mu + 1)\ Z_s}{R_p + Z_s} = \frac{(48 + 1)\ (8,900)}{(13,300 + 8,900)} = 19.7$$

where R_p and μ are the plate resistance and amplification factor of the 5876 pencil triode operating with 150 volts direct plate voltage and 5 milliamperes plate current. The plate current is kept low to reduce power requirements for the amplifier.

The input resistance is given by

$$R_{in} = \frac{R_p + Z_s}{\mu + 1} = \frac{13,300 + 8,900}{48 + 1} = 453 \text{ ohms}$$

The stage voltage gain is the tube gain multiplied by the step-down ratio of the plate cavity and the step-up ratio of the input cavity.

$$\text{Stage Gain} = G\sqrt{\frac{Z_x}{Z_s}\frac{R_{in}}{Z_o}} = 22.8\sqrt{\frac{300}{8,900}\frac{453}{50}} = 10.9$$

where Z_x is the impedance of the crystal and Z_0 is the characteristic impedance of the input. All coupling circuits are assumed to be perfect impedance matches. In a practical case gain of this magnitude cannot be realized.

<u>Crystal Mixer</u>

Crystal rectifiers are superior to triode vacuum tubes as frequency converters because of the relatively small amount of noise generated. This is especially true at frequencies above 500 megacycles. Crystals have less conversion loss at higher frequencies

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[8] Jones, M. C., "Grounded Grid Radio Frequency Voltage Amplifiers," <u>Proceedings of IRE</u>, Vol. 32, No. 7, July 1944, Page 423.

than diodes because the transit time effects are negligible.

A crystal rectifier consists of a crystal of silicon, germanium, or other semiconducting material and a tungsten probe that touches the surface of the crystal. The whole assembly is mounted in a small cylindrical container called the crystal cartridge. The junction of the probe and crystal show a nonlinear resistance to the flow of current. In fact the ratio of forward to back resistance causes the crystal to rectify alternating current. For small voltages the forward resistance of the crystal diode decreases with increasing voltage but levels off to a constant value as the voltage becomes larger. This constant resistance is the "spreading resistance" R_S, which is the resistance to current entering the silicon from the tungsten. It is in series with the nonlinear junction resistance r_b, which is shunted by the junction capacity, C. This capacity is in the order of one micromicrofarad and acts as a radio-frequency shunt. At high frequencies it will carry reactive current around the nonlinear resistance r_b, and reduce its

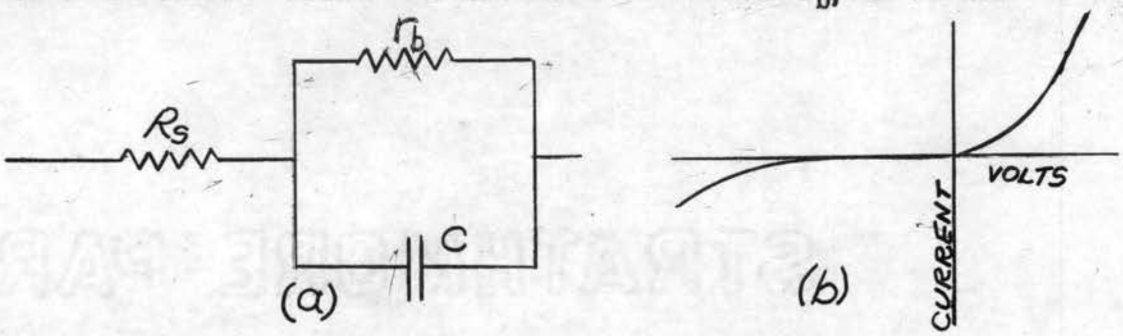

Figure 9. Equivalent circuit and diode characteristic of a typical crystal.

rectification efficiency. The capacity is reduced by reducing the size of the junction but this increases the danger of crystal burnout. The type of crystal used in the converter is the 1N21B.

It was chosen because it is a commonly used type, making it easier to obtain replacements. Its shunt capacity is only 0.3 micromicrofarads.

The action of the crystal is to combine the radio-frequency signal and the local-oscillator signal to give a sum and difference frequency. This is a characteristic of mixing action in any nonlinear circuit. In most superheterodyne converters the difference frequency is used and for this converter the difference frequency is 30 megacycles. The input impedance to the radio-frequency signals is approximately 300 ohms and the output impedance is approximately 400 ohms. The conversion loss, noise temperature, and impedances are all affected by the local-oscillator power injected into the crystal. With very little local-oscillator power injection the conversion loss is high. At high values of local-oscillator power injection it is again high because of the increase in back resistance. The best operating range for conversion loss and noise figure for a typical crystal is where the rectified current is between 0.3 and 2 milliamperes. The operating impedances do not change appreciably with oscillator injection unless the frequency is very high.

The crystal cartridge, X101, is contained in a crystal mount which is inserted into the side of the radio-frequency amplifier. The crystal mount is a small cylinder which is lined with teflon insulation to insulate the crystal cartridge from ground. The tip of the crystal is inserted in a small receptacle which in turn is connected to the output coupling loop of the amplifier. The top connector of the cartridge is gripped by a set of contacting fingers which is electrically connected to a small coaxial cable leading to the intermediate-frequency amplifier. The crystal has a cap, which screws on, to retain the crystal and add a small amount of filter capacity for radio frequency. As previously stated the local-oscillator signal is injected into the input circuit of the radio-frequency amplifier and is coupled out of the amplifier to the crystal mixer along with the signal frequency. The intermediate-frequency signal which is cabled to the intermediate-frequency amplifier is transformer coupled to the first stage of the amplifier. The grounded end of the transformer winding may be opened and used to measure crystal current. The current for the crystal mixer of the converter unit is approximately one milliampere. Practically all of this current is a result of local-oscillator injection.

Preselector

The preselector of the converter is essentially a resonant circuit consisting of a section of short-circuited coaxial transmission line tuned with a variable capacitor. A cross-section view of the preselector is shown in Figure 7. The variable capacitor, C101, is made of a cup-shaped plate attached to the center conductor of the transmission line and a cylindrical projection on the end plate of the preselector. This capacitor is adjusted by moving the center conductor back and forth by means of its threaded end. The short-circuited section of transmission line is less than a quarter-wavelength long and acts as an inductive reactance as discussed previously. This reactance is tuned to resonance with the capacitive reactance of the capacitor at the incoming signal frequency of 600 megacycles. The effect of adjusting the center conductor is twofold in that the line length and the capacity change simultaneously.

The preselector uses 0.875 tubing and 0.250 inch rod for its two conductors. The 70 ohm characteristic impedance is consistent with the design of the other units and is a high Q circuit. The high loaded Q of the preselector makes it useful for several reasons. It greatly attenuates image frequencies thereby offering a high degree of image rejection. It will attenuate other signals outside the passband. Transmitter spill-through and receiver blocking is greatly reduced. Some attenuation at the desired frequency will be introduced since it is a passive element. This is one of its disadvantages but measurements and calculations show the attenuation not to be serious.

The 3 decibel bandwidth of the preselector is the frequency difference (Δf) where the voltage is 3 decibels down (half power) from its resonant value. These measurements are made with a constant input voltage and constant input and output impedances. The bandwidth can be expressed as $\Delta f = f/Q$, where f is the center frequency. It is more practical when making measurements to find the value of Q in terms of f and Δf. The bandwidth should be more than the 3 megacycle bandwidth required by the overall receiver.

Energy is coupled into and out of the preselector by means of two magnetic coupling loops of 0.075 square inches cross sectional area. These loops are located 0.75 inches from the short-circuited end of the cavity. This distance is a compromise with the radio-frequency amplifier. The loops in the preselector should have been nearer the end where the current is maximum and the input tap for the radio-frequency amplifier should have been nearer the grid of the pencil triode. Since it was desired to keep the coupling loops directly opposite the input to the amplifier the intermediate location was chosen. The loops are of equal area and are spaced 150° apart about the periphery of the preselector. When the preselector is tuned to resonance the impedance seen at its input should be very near the same as that seen into the amplifier to which the preselector is coupled.[9]

The formula for the unloaded Q of the coaxial resonator has been given earlier in this chapter as

$$Q_o = \frac{0.694\sqrt{f}\ (b)\ \log (b/a)}{1 + b/a}$$

For the signal frequency of 600 megacycles this gives an unloaded Q of 2100 with no assumed losses in the dielectric or in the plates of the tuning capacitor. The Q for a loaded matched quarter-wavelength preselector may be expressed as

$$Q_L = \frac{\pi Z_0}{8} \frac{r^2 R_g}{A^2 \mu_o^2 f^2 \cos^2 \theta}$$

r = radius of loops from center of coaxial cavity.	0.008 meters
R_g= generator and load impedance.	50 ohms
A = area of coupling loops.	0.075 sq.in; .483 x 10^{-4} sq.met.
μ_o= permeability for air.	1.257 x 10^{-6} hy. per meter
θ = angular location of loop from short-circuited end.	13.7°; cos θ = 0.972

The loaded Q as calculated from this formula is 70 which gives a 3 decibel bandwidth of 8.6 megacycles.

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[9]Radio Research Laboratory Staff. Very High Frequency Techniques. Vol. II, McGraw-Hill 1947, Pages 773, 774.

The pass-band insertion loss due to dissipation is given by[10]

$$D = 20 \log \frac{Q_o}{Q_o - Q_L} = 0.28 \text{ decibels}$$

In other words, under ideal conditions the preselector passes about 97 percent of its input voltage to the radio-frequency amplifier.

The magnetic coupling loops are located 150 degrees apart instead of directly opposite each other as a matter of convenience in physical layout. The signal input connector could be made to occupy less space with the 150 degree angle. There is perhaps some direct coupling between the two loops as well as some coupling from the electric field in the cavity.

Converter Unit Development

The description of the various component parts of the converter has been presented essentially as it applies to the final model. As previously stated the components were first designed on an individual basis and later combined into a unit called the converter. As is characteristic of many projects of this type there is always a certain amount of laboratory investigation carried on in search of what appears to be the best solution to the problem. A brief resume of the difficulties and procedures of development will be given here but no attempt will be made to go into great detail.

The oscillator was the first unit which was designed and the one which required the most time in development. The first difficulty was in obtaining proper feedback for oscillation. The grid-cathode

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[10]Radio Research Laboratory Staff. Very High Frequency Techniques. Vol. II, McGraw-Hill 1947, Page 745.

cavity was too short. Various styles of slotted grid cylinders were built with various degrees of success in obtaining feedback for oscillation. The slotted grid cylinders allowed energy from the grid-plate cavity to radiate into the grid-cathode cavity to sustain oscillation. This scheme was abandoned in favor of a solid sleeve and dielectric lengthening of the grid-cathode cavity. The feedback probe was used in all cases as a means of adjustable feedback. The traveling dielectric plunger was the only method considered for frequency control. Metal contacting plungers were complications in the design and there was a tendency to avoid components with moving contacts. The first dielectric plungers used were made of Scotchcast, a plastic material manufactured by Minnesota Mining and Manufacturing Company. It has a dielectric constant of three at 600 megacycles which gave a frequency tuning range of 20 megacycles, but its mechanical properties were not suitable for high temperatures. It was replaced with Polypenco FM 10001 nylon which has electrical properties similar to that of Scotchcast. It appeared to be an answer to the dielectric plunger problem.

The amplifier presented no great problems. Its geometrical features are very much like those of the oscillator. It was necessary to find a means of lowering its center frequency of amplification from 625 megacycles to 600 megacycles without making the unit longer physically. This was done by adding a lumped capacity consisting of a thick brass disc about the plate-line anode connector. The dielectric plunger is the same as that used by the oscillator and gives the amplifier a frequency spread sufficient to track the oscillator.

The preselector is very simple in its design. The primary problem was to obtain an adjustment for the capacity which would not be critical with motion of the adjusting screw. This was done by shaping the plates so that the capacity change is more linear than it would have been with parallel-plane plates. At first appearance the preselector was a unit which would function without any difficulties.

The individual units were assembled into a "breadboard" converter and their performance was checked in conjunction with an intermediate-frequency amplifier. A makeshift crystal holder was inserted at the output loop of the radio-frequency amplifier. The input circuit of the intermediate-frequency amplifier had to be modified to match the low output impedance of the crystal. The intermediate-frequency amplifier formerly had a high impedance input. The first measurements of overall receiver sensitivity showed a requirement of 12 microvolts for a detected pulse signal which was twice the voltage level of the noise voltage. This is more briefly referred to as 12 microvolts for 2/1 signal/noise ratio. This measurement was made with an intermediate-frequency bandwidth of 2.5 megacycles.

No extensive tests and measurements were made on the "breadboard" unit operating as a converter. It was felt that a converter built as a single unit would have problems of its own and any time spent in working out design problems would more advantageously be spent on the combination-model converter. The "breadboard" model had served its purpose in that it showed that a converter unit of this type was feasible.

The first combination-model used units very similar to the "breadboard" model except they were arranged in a cluster as shown in Figure 10. The coupling from one unit to another is indicated in Figure 7. A few minor modifications of the components was necessary for the converter to function properly. The oscillator output coupling loop was reduced in size to reduce the amount of local-oscillator signal injected into the crystal. The oscillator and amplifier frequency tuning range had to be altered slightly. They were different because of differences in the bypass capacities at the short-circuited ends of the cavities. The general performance of the converter was reasonably good. The 2/1 signal/noise sensitivity ranged between 10 and 20 microvolts over the frequency band of 590 to 610 megacycles.

Two additional models were built which were very similar to the first combination-model. The major difference was that they were cadmium plated. This is the finish which is given to all brass parts in the laboratory. The two units were kept on an equal status at all times to see if their performance was identical. They were considered as final models and were used in obtaining all performance data. The sensitivity figure and the performance in general was much like the previous model.

It was with these models that it was found the nylon plungers were unsatisfactory. This trouble appeared during heat runs to check the frequency drift of the oscillator. The losses in the nylon began to increase immediately upon increasing the temperature and in some cases the oscillator practically ceased to function. The nylon was replaced with Polypenco Q200.5 which is the best known

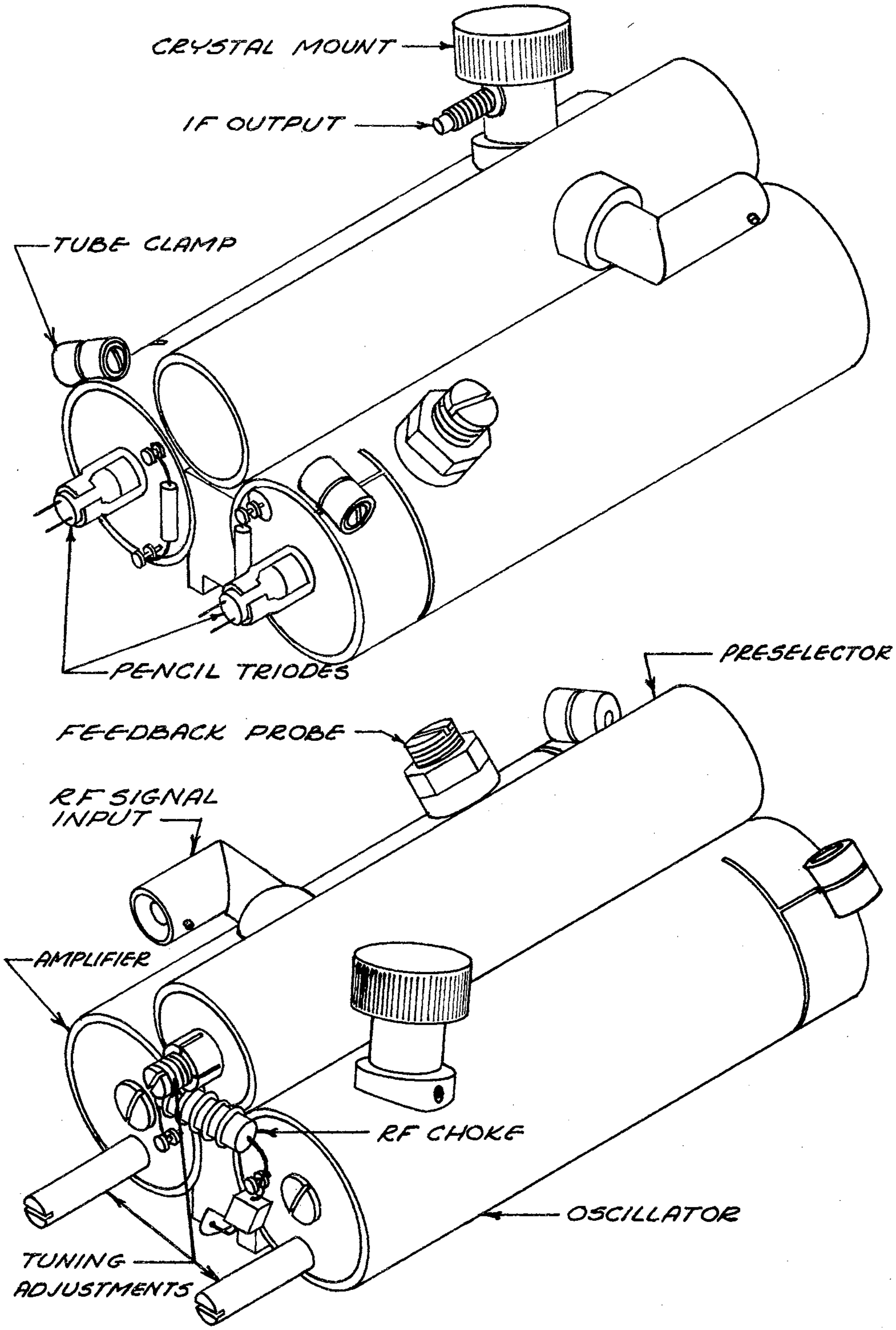

Figure 10. Two views of converter unit.

dielectric material to use for this purpose. Its dielectric constant is about 2.5 at 600 megacycles and its loss factor appears unaffected by temperature. The frequency spread of the oscillator was reduced to 16 megacycles with the new plunger material.

After the tuning plunger material problem was corrected the temperature runs on the oscillator were continued. It was found that at temperatures between 75°F and 225°F the units did but little frequency changing. They always had a drift range of less than one megacycle on various sets of data. When the ambient temperature was reduced below 75°F it was found that there were large frequency changes. One unit showed more pronounced changes than the other. It was observed that the break in the frequency shift curve was at room temperature and this fact was used in locating the trouble. It was found that the teflon spacer ring which supports the plate line within the grid cylinder was contracting at reduced temperatures. This contraction allowed the plate line to move in any direction in which residual radial forces existed. The capacity between the anode of the tube and the feedback probe has a pronounced affect on oscillator frequency as well as on feedback. Any change in this capacity due to movement of the tube and plate line will cause frequency drift. The teflon material has a coefficient of expansion of 10×10^{-5} in./in./°C. Since the teflon ring was fitted at room temperature the plate line and grid cylinder maintained concentricity at higher temperatures and consequently the frequency was more stable. To correct the trouble, the teflon rings were made with a taper on the diameter. The rings were inserted in the grid cylinder with the small end first. When the large end is in the cylinder the ring is under pressure. The frequency drift with the new spacer ring fell within specification limits of plus or minus one megacycle over a temperature range of -65°F to 225°F. This is a wider range of temperatures than the specifications require. A typical frequency-temperature curve is shown in Figure 11. Curves of this type usually do not repeat themselves precisely because of mechanical hysteresis, stability of measuring equipment, and stabilization time allowed for each reading.

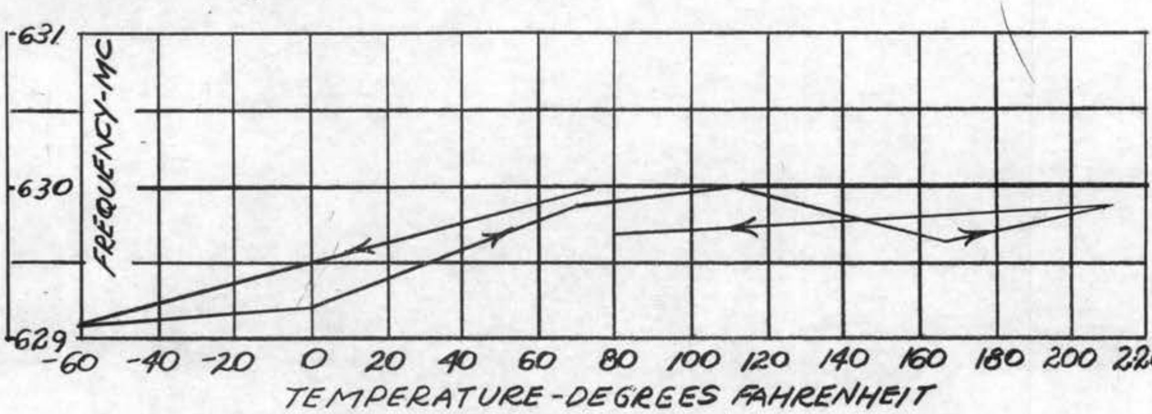

Figure 11. Typical plot of frequency vs. temperature.

Another problem which was encountered and which appeared to be serious was one of bandwidth. Since the 3 decibel bandwidth of the intermediate-frequency amplifier is three megacycles, the converter bandwidth would need to be more than three megacycles to guarantee an overall bandwidth of near three megacycles. It was felt in the earlier stages of development that the preselector and amplifier bandwidth would be sufficiently wide to not affect the overall receiver bandwidth. It was later found that both the preselector and amplifier had bandwidths of less than three megacycles and the overall receiver had a bandwidth of one megacycle when the preselector and amplifier were synchronously tuned. The bandwidth was widened by making the area of the coupling loops in the preselector and amplifier 33 percent larger. Their original widths were 0.38 inches and they were changed to 0.50 inches. The formula for Q on page 30 shows that the bandwidth which is $\Delta f/Q$, is proportional to the square of the area of the coupling loops. The 3 decibel bandwidth of all units is shown with the passband curves of Figure 12. The selectivity of the converter suffered considerably by lowering its Q. The overall receiver sensitivity remained about the same being about 15 microvolts for 2/1 signal/noise ratio.

Test Equipment and Performance Data

The test set-up generally used for most work is shown in block diagram form in Figure 14. The radio-frequency Signal Generator is a Measurements Corporation Model 84 or a Hewlett Packard Model 612A. The Sweeping Oscillator is a Kay Electric Model II. The Synchroscope is a Dumont 256D and the Oscilloscope is a Dumont 241. Other minor pieces of equipment such as meters are not mentioned. An All American Vibration Testing Machine is available for vibration testing. It can produce vibration frequencies over a range of 10 to 60 cycles per second with double amplitudes from 0 to 0.25 inches. It will produce a maximum acceleration of 1000 divided by the weight of the object being tested. The shock testing machine is a 60 pound pendulum which is allowed to swing and strike a large concrete slab. It is equipped with bumpers which give it 15 g acceleration for a duration of 0.011 seconds. The temperature chamber is capable of producing an ambient temperature range between -65°F and 250°F.

Most of the performance data has been mentioned previously but it will be summarized here. The figures given below are typical and may be applied to either or both of the cadmium plated models.

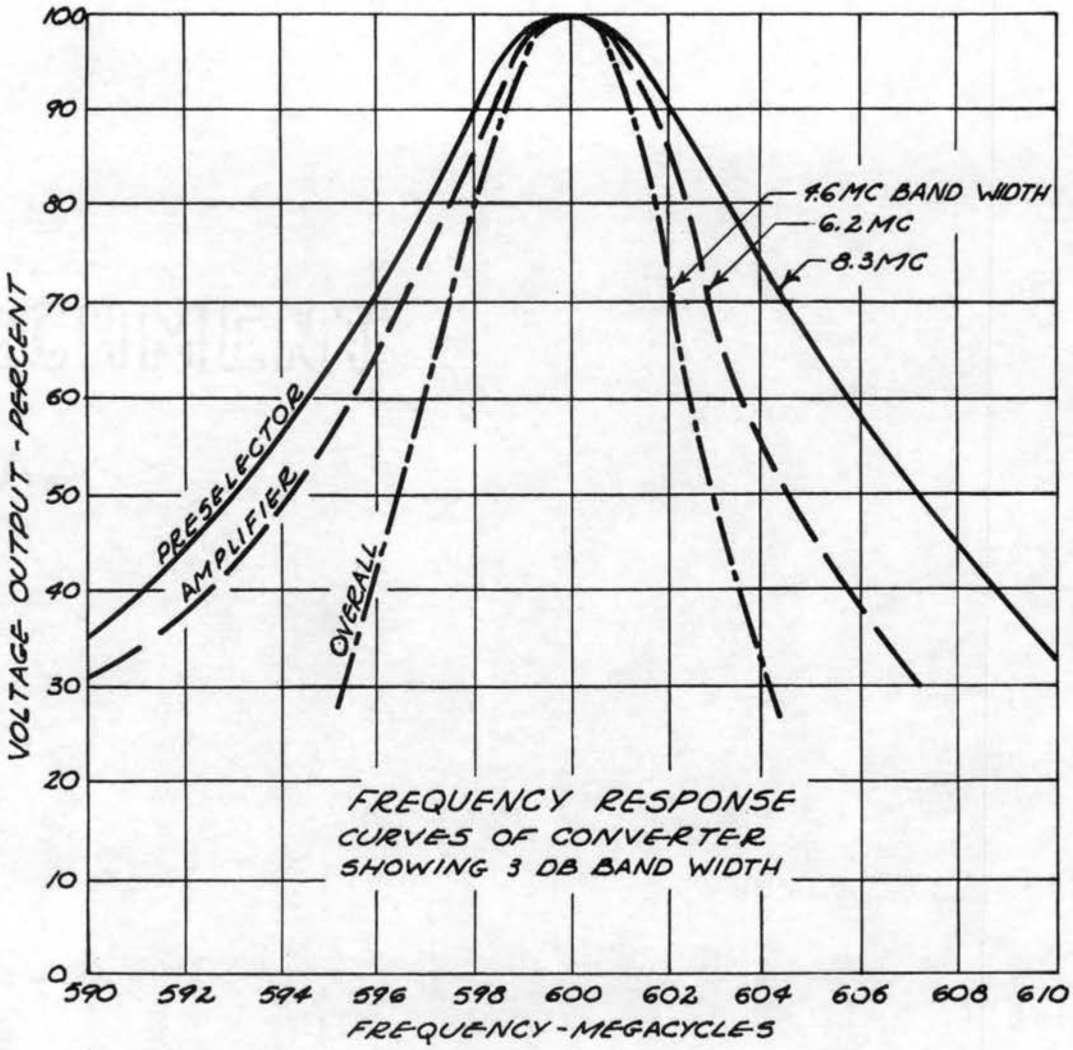

Figure 12. Frequency response curves of converter.

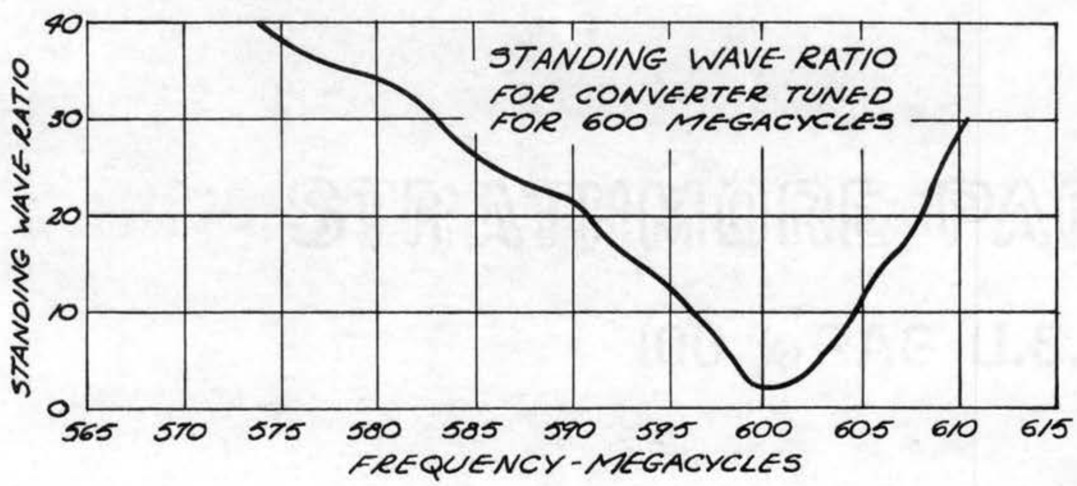

Figure 13, Input standing wave ratio.

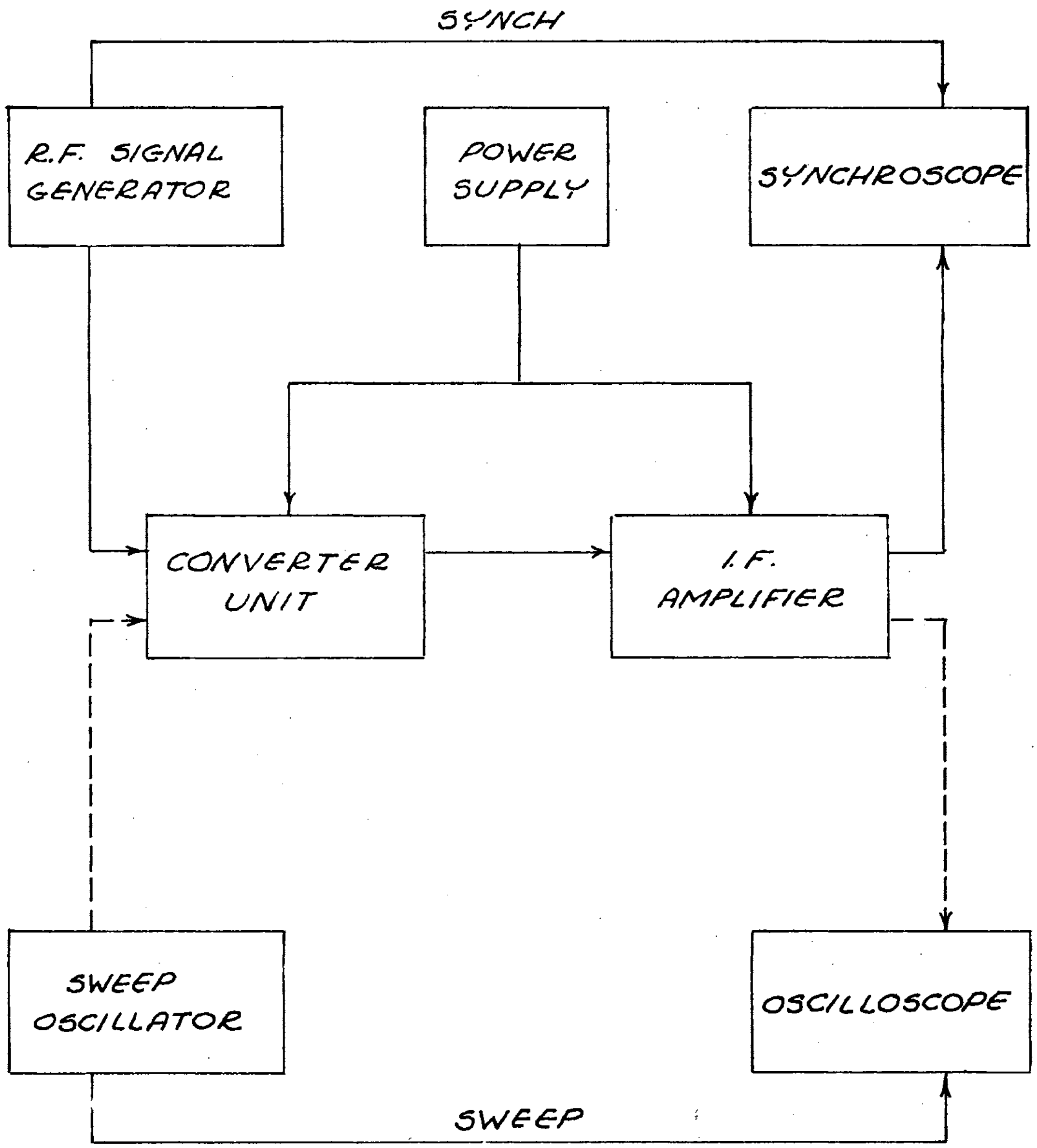

Figure 14. Block diagram of test set-up.

The intermediate-frequency amplifier used in making some of the measurements is one similar to the one with which the converter will be used. It has a 3 decibel bandwidth of three megacycles centered at 30 megacycles, and a gain of approximately 80 decibels.

The converter and the intermediate-frequency amplifier have a 2/1 signal/noise sensitivity of 15 microvolts with an output noise level of 0.05 volts from the video detector. Forty microvolts will produce a 0.5 volt video output pulse from the video detector of the intermediate-frequency amplifier. The converter has an image rejection voltage ratio of 158/1 or 44 decibels. The tuning range is 592 to 608 megacycles. The pass-band information is shown in Figure 12. It can be seen that the bandwidth of each unit gets progressively smaller from the preselector to the intermediate-frequency amplifier. Actual measurements on the preselector indicate a bandwidth of 8.35 megacycles and an insertion loss of 0.92 decibels. The calculated bandwidth is 8.6 megacycles whereas the calculated insertion loss is 0.28 decibels. The bandwidth of the amplifier is 6.2 megacycles and the overall bandwidth of the converter is 4.6 megacycles.

Mechanical testing of the converter has shown no detectable faults in the performance. The unit was mounted upon the All American Vibration Testing Machine and was operating with an intermediate-frequency amplifier during vibration. The converter was subjected to three, three-minute scans of vibration between 10 and 60 cycles per second in each of its three major planes. After an initial repair of a loose preselector tuning adjustment the unit performed satisfactorily under vibration conditions. The converter was also subjected to three impact shocks in both directions of its three major axes. The impact shocks were of 15 g acceleration and had a duration of 0.011 seconds. The unit was non-operating during the shock test but performance data before and after the test indicated no change in performance.

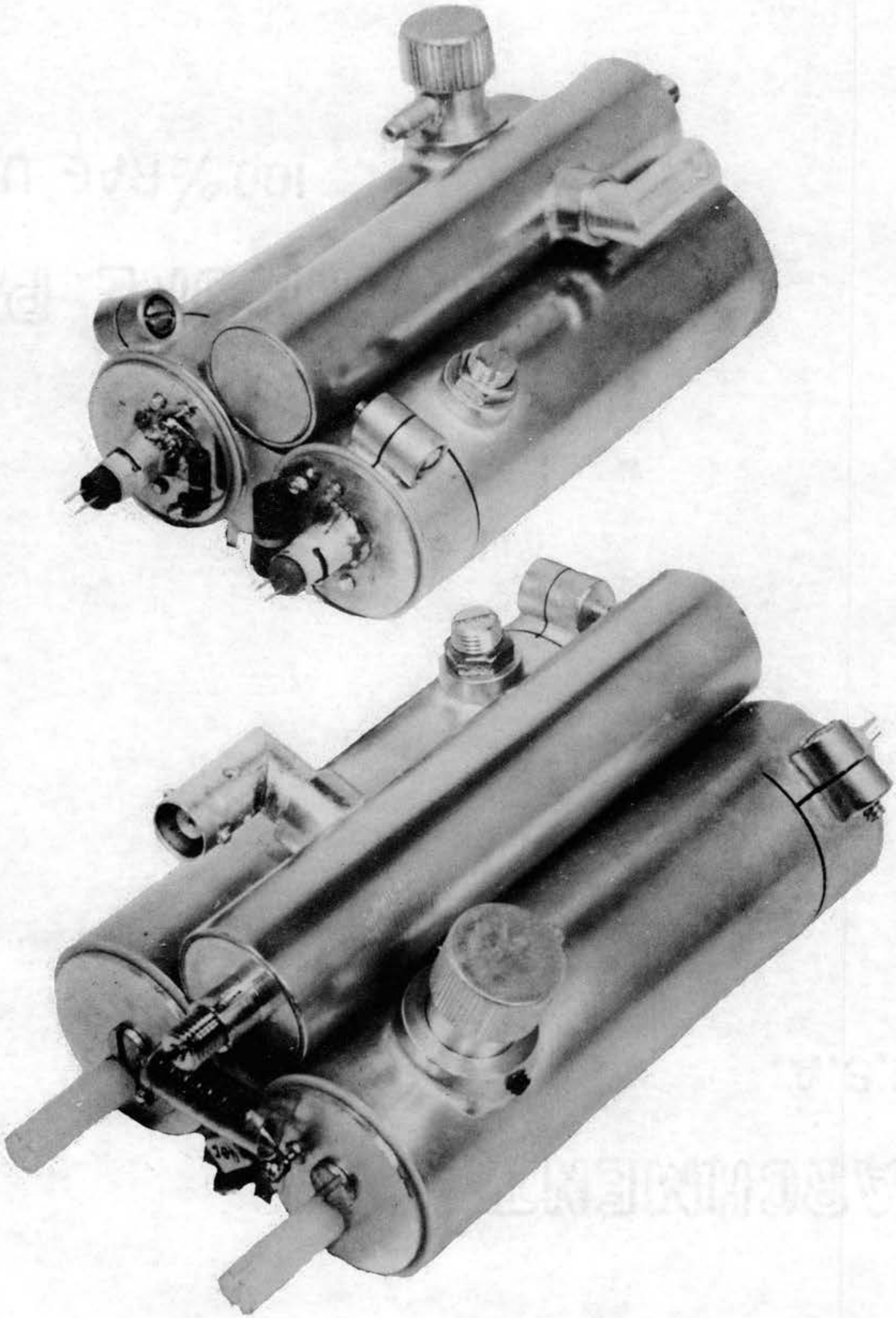

Figure 15. Photograph of converter.

CHAPTER IV

SUMMARY AND CONCLUSIONS

The design and development of this converter has produced a unit which is in many ways superior to its predecessor. It is felt that the general performance, reliability, and ease of reproduction of the new unit is better than the old converter. Below is tabulated several comparative figures for the two units which show desirable and undesirable features of each.

	Old Unit	New Unit
Outline dimensions - inches	2 x 3 x 4	2 x 2 3/4 x 4 3/4
Weight - pounds	0.75	1.5
Plate current at 150 volts - milliamps.	23	13
Filament current at 6.3 volts - milliamps.	400	270
Sensitivity for 2/1 signal/noise - microvolts	35	15
Image rejection ratio	3/1	158/1
Maximum frequency-temperature drift - mc.	1.5	1
Frequency tuning range - mc.	580-620	592-608

All the above figures, with the exception of the weight, favor the new converter. The cost of production may be somewhat greater for the new converter but it is felt that this will be offset by its reliability.

If the converter units continue to perform satisfactorily with production quantities, their use will probably be continued through any future beacon orders. To date the converter is being designed into a new developmental model beacon for the U. S. Air Force. A small production quantity of eight units is also being built for a beacon order which originally used the old converter.

Any future development or improvements on the converter should tend toward a reduction in weight and improvement in signal sensitivity. It is feasible to reduce the weight by using brass tubing with a smaller wall thickness or by using a lighter weight metal than brass.

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